The two-inductor "T" impedance matching network

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THE matching networks required to couple high frequency power amplifiers to their loads are usually in the form of either a "TT" or a "T", both realized with two capacitors and one inductor. An alternative "T" network using two inductors and one capacitor appears to have significant advantages. The network is especially appropriate to the design of lumped constant-impedance matching networks for vhf power amplifiers using either transistor vmosfet or valve. It is also appropriate to hf use where it may have advantages over the single "TT" network when the high shunt capacitance consequent upon parallel operation of multiple output devices causes design problems.

Although superficially a single "T" network, the derivation shows it to be a two-stage transformation in which elements of the two parts are lumped into a single inductor. This results in unusually high values of inductance relative to the design frequency and impedances, enabling considerable lead inductance to be tolerated even at vhf.

Considered only as impedance transformers, all the possible three-element narrow-band impedance matching networks (Fig 1) are equally suitable. When practical considerations such as available or realizable component values, unavoidable stray reactances, circulating currents, applied potentials and the requirement for adjustment and dc blocking are added, it is apparent that all networks are not equally suitable.

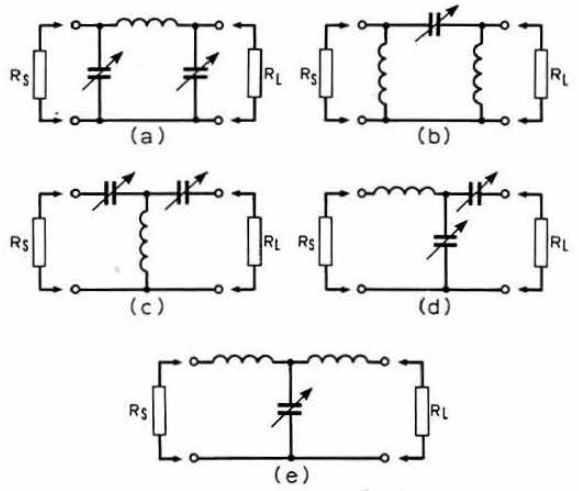

Fig 1. Five three-element matching networks

High-power devices and their associated cooling mechanisms are invariably relatively large. A natural consequence is a necessity for connecting leads with a self-inductance which may have quite serious effects at higher frequencies. Similarly, large devices, especially when parallel connected or coupled to a heatsink, can have an output shunt capacitance much greater than the desired input capacitance of a correctly designed "TT" section.

At full amateur power ratings the circulating current in the network may be as high as 20A, and any capacitors in its path must be able to carry it without excessive loss. This makes high-power transmitter variable capacitors of more than a few tens of picofarads physically large, and also gives them appreciable self-inductance. If it is necessary to have variable capacitors in a network, then networks which allow one terminal of the capacitors to be grounded will be preferred.

The five three-element matching networks (Fig 1)

(a) This two-capacitor "pi", much used in hf power amplifiers, has a range of impedance ratio and "Q" which is limited by the shunt capacitance of the amplifier device. This renders it unsuitable for vhf use, and may even pose problems at 30MHz when high input impedances are required.

(b) This two-inductor "TT" is theoretically less limited. The capacitor element, which is the natural choice for adjustment, has both ends "hot" to rf and, unfortunately, the section is "high-pass" and therefore provides no harmonic attenuation.

(c) This two-capacitor "TT" is also a high-pass function with both capacitor terminals "hot".

(d) This two-capacitor "T" network is well suited to low impedance sources, hence its popularity in transistor power amplifiers. The source impedance cannot exceed the load impedance with this network, and both terminals of the series capacitor are "hot".

(e) This two-inductor "T" will match a very wide range of impedances at any practical "Q" factor. It can accommodate low shunt capacitive reactance at its input, and considerable lead inductance can be absorbed into the two series inductors. At vhf, even when coupling from high source (1kΩ) to low (50Ω) loads, the capacitor is satisfactorily low in value and it has the required one terminal at ground. It is also a "low-pass" function giving good harmonic attenuation.

Derivation

It is instructive to follow how the two-inductor "T" is derived stage by stage; the objective is to transform the load, which will usually be a pure resistance of 50 or 75Ω, to the desired load resistance for the amplifier. The amplifier usually shunts the transformed load resistance with its output capacitance. At higher frequencies this shunt reactance may be more than one order lower than the optimum load for the amplifier, and if not dealt with will effectively short-circuit the amplifier.

Starting from the amplifier end, the network is terminated by the required transformed load impedance shunted by the device output capacitance (Fig 2). This complex termination is

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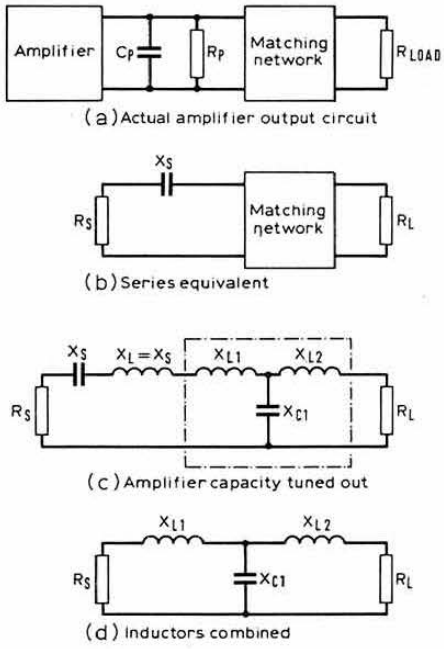

Fig 2. Evolution of the two-inductor "T" network

first turned into its series equivalent. (Many current high frequency power semiconductor data sheets give this impedance in the series form as Z_{OL} or Z_{out}.)

Having obtained the equivalent series reactance, an equal reactance of the opposite kind is connected in series, eg a capacitive reactance requires a numerically equal inductive reactance. The total series combination is thus effectively reduced to a pure series resistance.

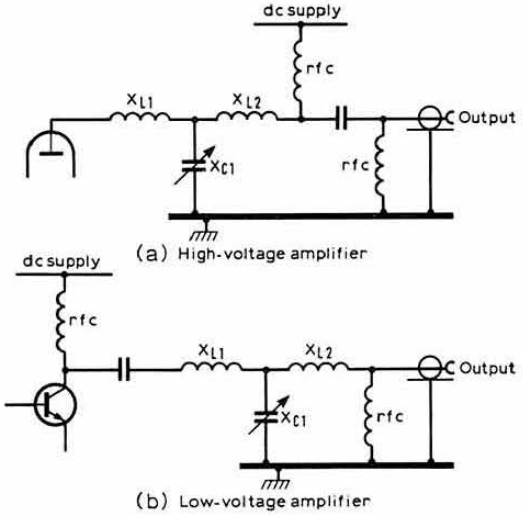

Fig 3. DC supply feed to the network

An impedance transformer is now required to transform the load to the amplifier series resistance. By choosing the two-inductor "T" which is the dual of the more familiar "TT", the series tuning inductor may be combined with one inductor of the "T", reducing the network to three elements.

The actual procedure for computing the values is as follows. If only the parallel equivalent of the amplifier load is known convert it to the series equivalent:

$$Rs = \frac{Rp\ Xp^2}{Rp^2 + Xp^2}$$

$$Xs = \frac{Rp^2\ Xp}{Rp^2 + Xp^2}$$

Select a value for "Q" and now calculate:

$$XL1 = Xs + QRs$$

$$XL2 = RL \sqrt{\left[\frac{Rs\,(1+Q^2)}{RL}\right] - 1}$$

$$XC1 = \frac{Rs\,(1+Q^2)}{Q + \sqrt{\left[\frac{Rs\,(1+Q^2)}{RL}\right] - 1}}$$

Hence find L1, L2 and C1.

When designing L1 and L2 allow for lead inductance.

Construction hints

The two inductors, and any chokes used to feed dc to the system, must be mounted with their axes mutually perpendicular to avoid mutual coupling.

DC blocking may be applied at either end of either inductor. The blocking capacitor may either have negligible reactance at design frequency, which is the usual practice, or it may be lumped with Xs, and XL1 compensated accordingly. Note that if the blocking capacitor is low in value, high rf potential may exist across it.

There will be a voltage node somewhere on the inductors, and this is the ideal dc feed-point. However, for low-voltage amplifiers a choke feed to the amplifier end will be satisfactory. For high-voltage amplifiers, blocking and choke feed at the load end will be more satisfactory (Fig 3).

Use thick wire, tube or strip for the inductors to carry the high circulating currents. Keep the capacitor leads short and direct.

A "T" network constructed on the above principles to match a 50Ω load to a 144MHz linear amplifier using a TY2-125 valve requiring a 4kΩ load and having a shunt capacitance of 6pF (X_P = 183Ω at 145MHz), gave an indicated efficiency of 63 per cent dc anode power to rf in the load under maximum single-tone condition. This is very close to the theoretical anode efficiency of the valve under the test condition, indicating a very efficient transfer of power to the load. A 5in anode lead required to mount the network well clear of the "big glass bottle" was absorbed easily into XL1 (294nH) contributing about half the required inductance. XC1 came to about 14pF. □